Sensorless Vector Control of Induction Motors

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***Abstract* – In this paper, two structures based on the partial induction motor model and optimization procedures for speed and flux rotor of the motor are overviewed. Simulation researches of sensorless direct vector control of induction motors using the proposed estimators are accomplished. The researches demonstrate their sufficient performance during motor parameter (rotor resistance) variations, disturbance and wide range variable references input control signals.**

***Keywords* – Sensorless vector control, Induction motor, Flux estimation, Speed estimation**.

I. INTRODUCTION

The interest for sensorless induction motor drives has been constantly rising during the last decade. They fill the middle ground between high-performance closed-loop control and simpler open-loop (V/Hz) control of induction motor (IM). The advantages of using these systems are: reduced hardware complexity, reduced size, no sensor cable, increased reliability, less maintenance requirements, lower cost, better noise immunity. Most sensorless AC drives are based on flux vector methods, hence loosely called sensorless vector (SV) control. There are many models of sensorless speed controllers described in the literature based on the extended Kalman filter theory [1,2], linear state Luenberger observer [2,3], neural network [3] and others [4-7]. Basic problem in the area of sensorless control is the estimation of the low speed. Here two structures based on the partial induction motor model and optimization procedures for speed estimation and for flux rotor of the motor, are proposed.

The paper is organized as follows: Section II describes the induction motor model, in Section III are described the proposed structure estimators, Section IV investigates the simulation testing of sensorless vector drive, and Section V summarizes the conclusion about results.

II. INDUCTION MOTOR MODEL

The equations for electrical equilibrium of voltages of the IM with cage rotor in the stationary α – β frame, introduced in a complex form are:

$$\bar{u}_s = R_s\bar{i}_s + \frac{d\overline{\Psi}_s}{dt} \tag{1}$$

$$0 = R_r\bar{i}_r + \frac{d\overline{\Psi}_r}{dt} - jp_p\omega_r\overline{\Psi}_r \tag{2}$$

$$\overline{\Psi}_s = L_s\bar{i}_s + L_m\bar{i}_r \tag{3}$$

$$\overline{\Psi}_r = L_m\bar{i}_s + L_r\bar{i}_r \tag{4}$$

where: $\bar{u}_s, \bar{i}_s, \overline{\Psi}_s, \bar{i}_r, \overline{\Psi}_r$ are representation vectors of the stator variables (voltage, current and flux) and rotor variables (current and flux); ω_r – rotor angular speed; R_s, R_r - stator and rotor phase resistance; L_s, L_r stator and rotor phase inductance; L_m – mutual inductance; p_p – number of the pair poles. For each vector $\bar{x}$ (where x means voltage, current, flux) is valid $\bar{x} = x_\alpha + jx_\beta$ where x_α, x_β are representation vector projections on α and β axes.

After elimination of $\overline{\Psi}_s$ and $\bar{i}_r$ from (1÷4), it is obtained:

$$\bar{u}_s = R_e(T_e p+1)\bar{i}_s - \frac{R_r L_m}{L_r^2}\overline{\Psi}_r + j\frac{L_m}{L_r}p_p\omega_r\overline{\Psi}_r \tag{5}$$

$$\bar{i}_s = \frac{1}{L_m}(T_r p+1)\overline{\Psi}_r - j\frac{T_r}{L_m}p_p\omega_r\overline{\Psi}_r \tag{6}$$

where:

$$R_e = R_s + R_r\frac{L_m^2}{L_r^2};\ L_e = \frac{L_s L_r - L_m^2}{L_r};\ T_e = \frac{L_e}{R_e};\ T_r = \frac{L_r}{R_r};\ p = \frac{d}{dt}.$$

For the purpose of vector control system synthesis, electromechanical processes in the IM and the mechanism (for one-mass mechanical part) in orthogonal coordinate system d–q, orientated on the rotor flux vector are described by the following equations:

$$\begin{aligned}
u_{sd} &= R_e(T_e p+1)i_{sd} - L_e\omega_g i_{sq} - \frac{R_r L_m}{L_r^2}\Psi_r \\
u_{sq} &= R_e(T_e p+1)i_{sq} + L_e\omega_g i_{sd} + \frac{L_m}{L_2}p_p\omega_r\Psi_r \\
i_{sd} &= \frac{1}{L_m}(T_r p+1)\Psi_r \\
\omega_g &= p_p\omega_r + \frac{R_r L_m}{L_r}\frac{i_{sq}}{\Psi_r} \\
M_e &= k_m\Psi_r i_{sq} \\
M_e - M_c &= Jp\omega_r
\end{aligned} \tag{7}$$

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where: i_{1q}, i_{1d}, u_{1q}, u_{1d} - active and excite components of stator current and stator voltage, ω_g – co-ordinate system speed, M_e, M_c – electromagnetic and resistant moments, J - moment of inertia, $k_m=(3p_pL_m/2L_r)i_{sq}/\Psi_r$.

III. Speed end flux estimation

When designing a sensorless direct vector control of IM, there are a few considerable problems: determination of rotor speed, transitory position and magnitude of the support vector, which is most often the rotor flux vector. Two variants for dealing with this problem are proposed. They are based on IM partial model and the optimization procedure us.

A. Estimator 1

Estimator 1 is composed of two partial models of the motor (Model 1, Model 2) and optimization procedure (OP) – fig.1 (with thick line are shown vector variables and with thin line – scalar variables). Model 1 estimates the rotor flux $\hat{\bar{\Psi}}_r$ and forms $e^{j\theta}$. Model 2 gives estimation of the stator current $\hat{\bar{i}}_s$.

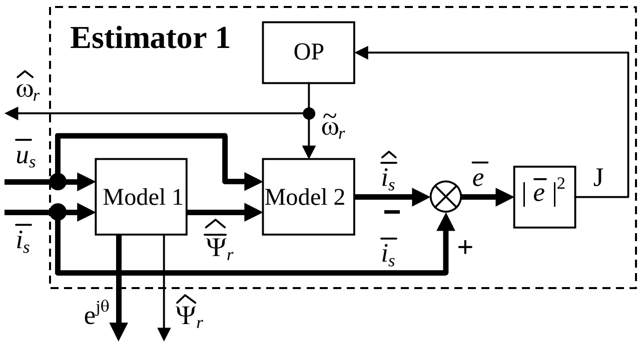

Fig. 1. Structural scheme of Estimator 1

Model 1 is described by the following equations:

$$\hat{\bar{\Psi}}_r = \frac{L_r}{L_m}\frac{\bar{u}_s - R_s\bar{i}_s}{p} - \frac{L_sL_r - L_m^2}{L_m}\bar{i}_s\,; \tag{8}$$

$$\hat{\Psi}_r = \left|\hat{\bar{\Psi}}_r\right| = \sqrt{\hat{\Psi}_{r\alpha}^2 + \hat{\Psi}_{r\beta}^2}\,; \tag{9}$$

$$\theta = arctg\frac{\hat{\Psi}_{r\beta}}{\hat{\Psi}_{r\alpha}}\,, \tag{10}$$

where θ is the angle between vector $\hat{\bar{\Psi}}_r$ and axis α.

The equation (8) is a result of (1) after expressing $\bar{\Psi}_s$ by trough $\bar{\Psi}_r$ and $\bar{i}_s$.

The estimated value of the stator current $\hat{\bar{i}}_s$ is obtained from Model 2 on the basis of equation (5), as for the purpose is used the measured stator voltage and obtained in Model 1 rotor flux:

$$\hat{\bar{i}}_s = \frac{\bar{u}_s + \frac{R_rL_m}{L_r^2}\hat{\bar{\Psi}}_r - j\frac{L_m}{L_r}p_p\omega_r\hat{\bar{\Psi}}_r}{R_e(T_ep+1)} \tag{11}$$

The Optimization criterion is:

$$\mathrm{J} = \left|\bar{e}\right|^2 = \left|\bar{i}_s - \hat{\bar{i}}_s\right|^2 \tag{12}$$

Through OP $\tilde{\omega}_r$ varies for each sampling period so that the criterion J be minimized and the estimation of the rotor speed $\hat{\omega}_r$ is obtained:

$$\hat{\omega}_r = \tilde{\omega}_r \text{ in case of } \mathrm{J} = \min \tag{13}$$

In the simplest case OP may be realized by method of consecutively searching in definite interval. For “k” tact $\tilde{\omega}_r(k)$ will be changed in the interval $\hat{\omega}_r(k-1) \pm \Delta\omega_{\max}$.

The determination of $\Delta\omega_{\max}$ is realized on the basis of following reasons. When working with constant flux (Ψ_r=const) and in the presence of current restriction, the maximum value of the electromagnetic torque is $M_{e\max} = k_m\Psi_rI_{sq\max}$, where $I_{sq\max}$ is the maximum permissible value of the active component of stator current with referent current restriction. The result for maximum dynamical torque is $M_{d\max} = \left|M_{e\max}\right| + \left|M_{c\max}\right| = \left|M_{e\max}\right| + M_n$, where M_n is nominal motor torque. From equation of motion (7) is obtained maximum possible change of speed:

$$\Delta\omega_{\max} = \frac{M_{d\max}T_0}{J_{\min}} \tag{14}$$

where T_0 – sampling period, $J_{\min}=J_m$, J_m – inertia moment of motor.

B. Estimator 2

Estimator 2 is composed by one partial model of IM (Model 3) and optimization procedure – fig. 2. The model gives simultaneously estimations for the rotor flux $\hat{\bar{\Psi}}_r$ and stator current $\hat{\bar{i}}_s$ and forms $e^{j\theta}$.

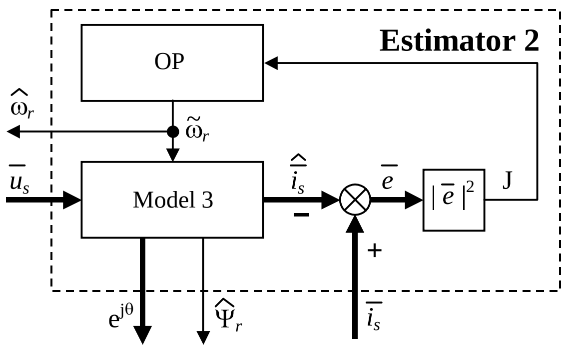

Fig. 2. Structural scheme of Estimator 2

Model 3 is described by equations (5) and (6). The optimization procedure is the same as this one, described in the first variant. Optimization criterion is defined by (12) and the estimation of the rotor speed $\hat{\omega}_r$ is obtained according to (13).

On the basis of the proposed speed and rotor flux estimators is obtained the direct vector control system of the IM – fig. 3. In the figure are used the following symbols: FC – flux controller; SC – speed controller; CCB and CCA – controllers of the active and excite components of the stator current; BC – block for compensation; abc/αβ, αβ/dq, dq/αβ, αβ/abc – co-ordinate transformers, PC – power converter, IM – induction motor, M – mechanism; ω_{ref}, Ψ_{ref} - referent values of speed and rotor flux; $\bar{u}^*_{dq}, \bar{u}^*_{\alpha\beta}, \bar{u}^*_{abc}$ - control signals.

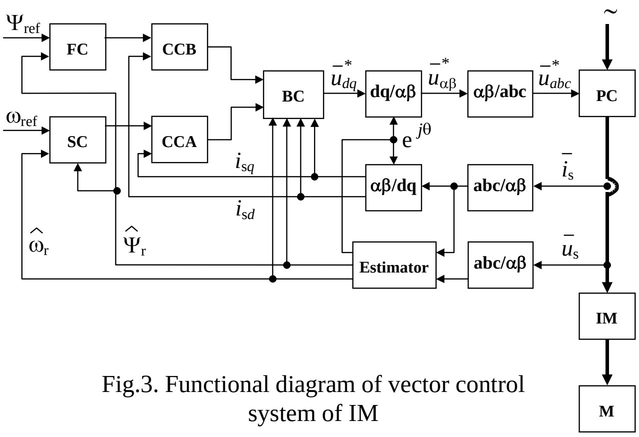

Fig.3. Functional diagram of vector control system of IM

IV. SIMULATION RESEARCH AND REZULTS

The proposed control algorithm has been tested using a 5 kW induction motor with rated torques 36.224Nm and nominal current I_{sn}=11.2A. In the simulation model coordinate transformations, non-linear and discrete PC properties are rendered in account. The researches are carried out using variable stator and rotor resistance, without/with noise in the input signals.

A part of the simulation research of the direct vector control system is shown on fig. 4÷7. On the fig.4 and fig.5 is represented the system performance using Estimator 1 and on the fig.6 and fig.7 - using Estimator 2.

Motor speed and rotor flux are compared with the ω_{ref} and Ψ_{ref} (fig.4a÷7a) and with the estimations $\hat{\omega}_r$ and $\hat{\Psi}_r$, derived from the estimators (fig. 4b,c,d,e÷7b,c,d,e). Fig.4,6 are obtained bay using the basis value of the motor electric parameters in absence of noise, fig.5,7 with R_s=1.2R_{skat} and R_r=1.2R_{rkat} (R_{skat}, R_{rkat} basis values of R_s и R_r) and in case of noise addition to the stator currents. The noise addition is simulated by additive white noise, ratio noise/signal=0.05. In all cases the estimators work with the basis values of the motor parameters.

The resistive moment M_c is accepted to be reactive and it changes in range of (0.1÷1)M_n as follows: M_c=0.1M_n for t≤0.2s; M_c=0.5M_n for 0.2<t≤1s; M_c=M_n for 1<t≤1.6s; M_c=0.5M_n for 1<t≤2s. Where differences between motor's resistances and basis values exist, a static fault comes out.

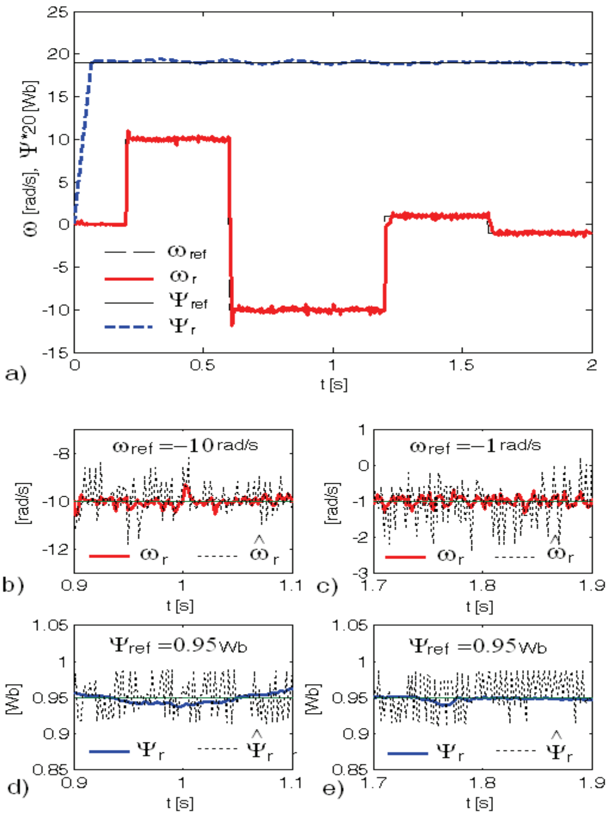

Fig.4. Low speeds in absence of noise and R_s=R_{skat} и R_r=R_{rkat} (with Estimator 1)

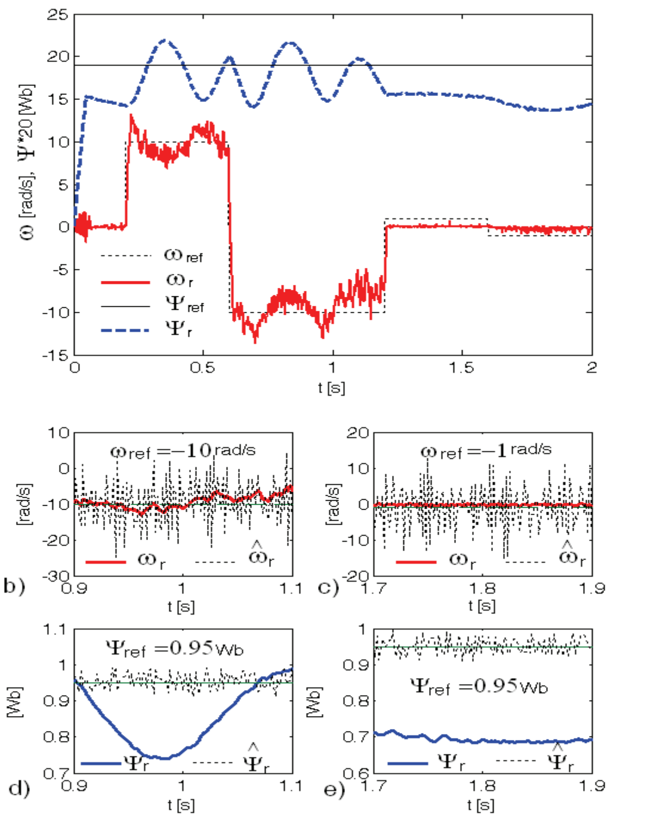

Fig.5. Low speeds in presence of noise and R_s=1.2R_{skat} и R_r=1.2R_{rkat} (with Estimator 1)

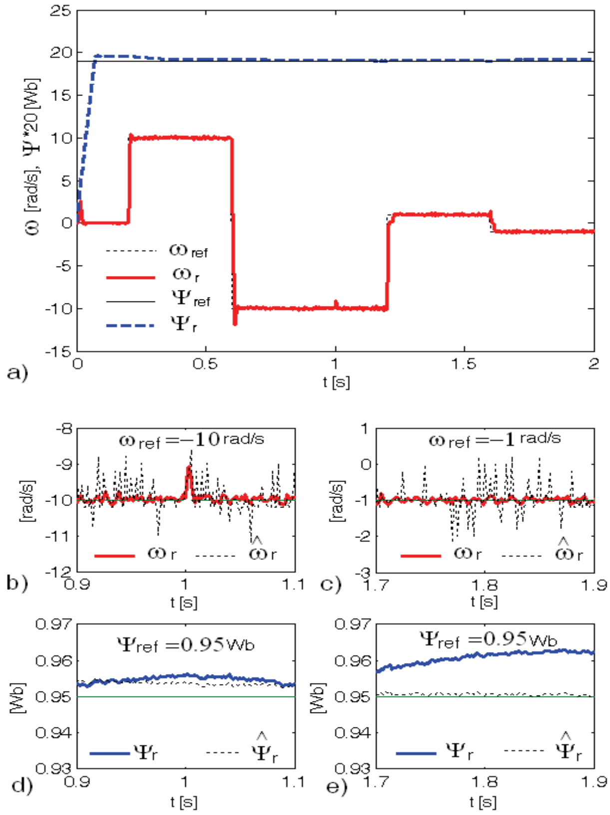

Fig.6. Low speeds in absence of noise and $R_s=R_{skat}$ и $R_r=R_{rkat}$ (with Estimator 2)

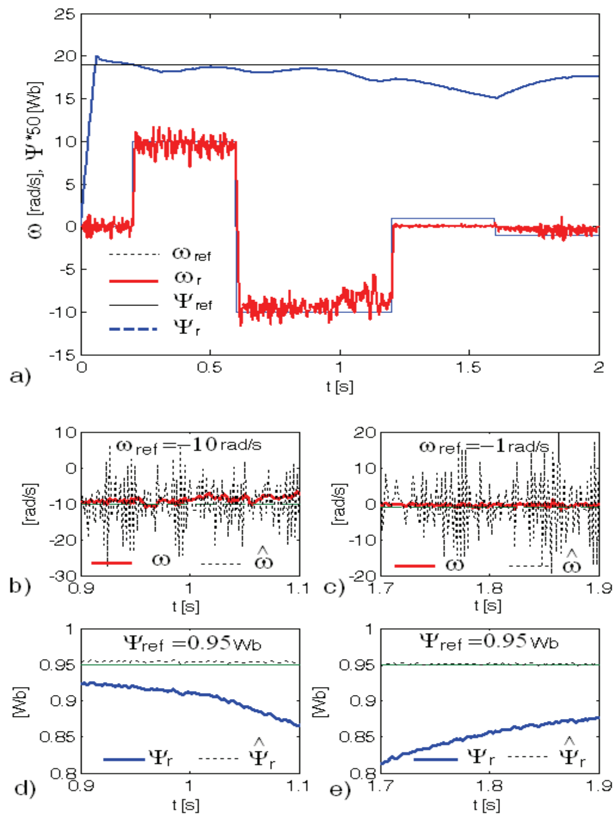

Fig.7. Low speeds in presence of noise and $R_s=1.2R_{skat}$ и $R_r=1.2R_{rkat}$ (with Estimator 2)

When there is difference between motor resistance and their basis values, a static fault occurs. This fault is proportional of the load and of the diversion of the real resistances from the basis values.

The researches accomplished confirm the capability of the system to work with both estimators. The second estimator variant is more robust regarding the parametric and signal disturbances. It makes a filtration of the estimated values of the rotor flux end for this reason it is actually a smooth curve. That is why the estimator from figure 2 is preferable.

V. Conclusion

Two structures using partial induction motor model and optimization procedures of speed and rotor flux estimation are proposed. On this basis two structures of direct vector control of IM drive are formed.

Simulation researches are done, confirming the efficiency of the proposed estimators and the vector control systems schemes. The research is carried through under the following conditions: wide range of speed control, variations of motor parameters (rotor end stator resistances), noise implemented in the signals of phase stator currents.

The researches demonstrate sufficient performance of sensorless IM drive control.

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